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TUNNELDIODES AS MULTIVIBRATOR FOR FAST COUNTING IN TIME-OF-FLIGHT CODING AND DERANDOMISING APPLICATIONS

by

H. MEYER and H. VERELST

1964

Joint Nuclear Research Center
Geel Establishment — Belgium

Central Nuclear Measurements Bureau (CNMB)

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A ten-stage scaler was built with a maximum rate greater than 300 Mc/s. With tunneldiode univibrators as coupling elements between stages the propagation delay is about 16 mμs with less than $5\,^{0}/_{00}$ jitter.

The use of such stages as "pulse derandomiser" is considered. The best measured double-pulse resolution was 2 mμs, but with the pulses momentarily available from fast nuclear counters, limits are mostly given by trigger pulse shape.

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TUNNELDIODES AS MULTIVIBRATOR FOR FAST COUNTING IN TIME-OF-FLIGHT CODING AND DERANDOMISING APPLICATIONS

1 — INTRODUCTION

In connection with a time-of-flight coding device under development(*), with a minimum channel width of one nanosecond, a fast address scaler with small and stable propagation delay should be realized. The equipment shall be used together with a planned universal data handling system for nuclear experiments, which is under study.

For accurate time determination, the propagation delay jitter of the scaler must be smaller than the minimum channel width, for normal temperature and supply voltage changes.
Transistor flip-flops using current switching, which have propagation delays of about 10 mμs per stage (1) are here not adequate.
Gated-counters (2) could have very short propagation delays (tunneldiode gates) but their double-pulse resolution is determined by the used transistor flip-flops and is normally worse than that of a tunneldiode multivibrator.
So a tunneldiode scaler was finally selected to have the possibility of working with a high counting rate. A small propagation delay, which also for a ten-stage scaler is only slightly longer and practically as stable as that of a gated counter, can be realized.

Collaterally the possible double-pulse resolution of the tunneldiode multivibrator was investigated in order to derandomize spectra with high rates. The use of some cascaded binaries with high double-pulse resolution, connected with a normal slow scaler, can avoid counting losses, by narrowing the pulse distribution around its mean value. For spectra with a great energy range the scale should allow a large amplitude range (dynamic).
In addition a high sensitivity is important to avoid the use of foregoing amplification.
As far as possible the device should be insensitive to pulse shape variations.
A definite trigger level would be useful.

2 — TUNNELDIODE MULTIVIBRATOR

The basic multivibrator circuit, as described in the literature (3 and 4) is represented in figure 1. The supply voltage U is restricted to a magnitude such that one tunneldiode is in the high voltage state and the other in the low voltage state.

The leading edge of the negative (positive) trigger pulse switches the tunneldiode, in the high (low) voltage state, from the high (low) state to the low (high) state. The trailing edge of the trigger pulse switches then the second tunneldiode to the high (low) voltage state, with the help of the memory function of the inductivity L.

2.1 — Equivalent circuit

In order to compute the fundamental dynamic behaviour of the circuit, the equivalent circuit represented in figure 2 is used and negative trigger pulses are chosen.

(*) See EUR 491.e

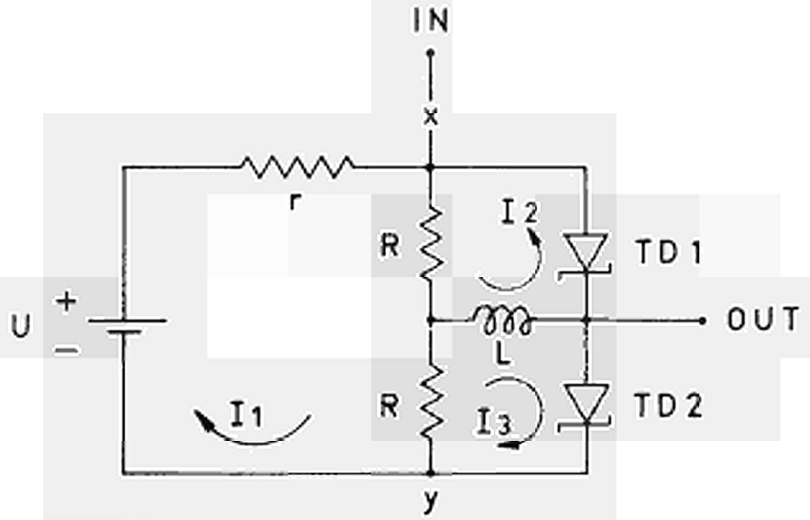

Fig. 1 — Schema of tunneldiode bistable multivibrator

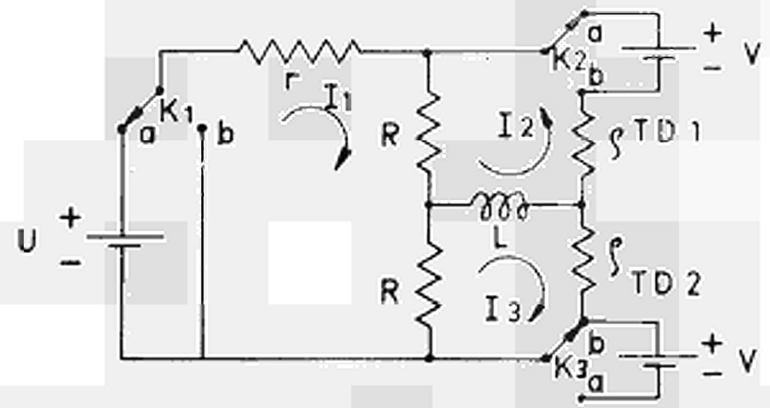

Fig. 2 — Equivalent circuit used for the computation of the qualitative behaviour of the flip-flop

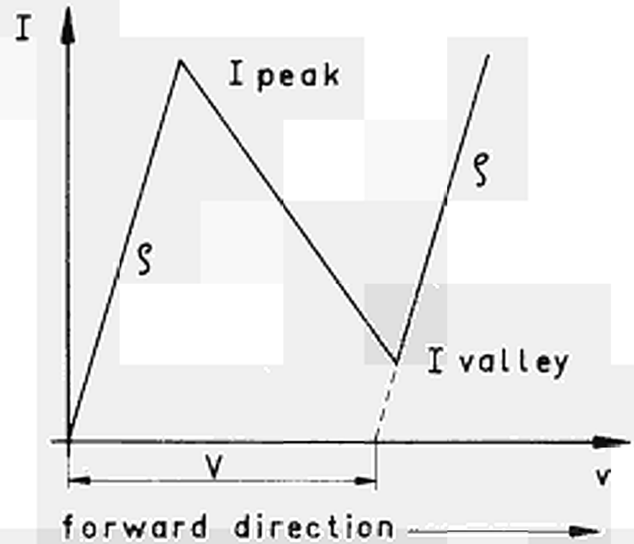

Fig. 3 — Simplified tunneldiode characteristic, as it was used in the circuit represented in fig. 2

Each tunneldiode is replaced here by an equivalent linear circuit consisting of an impedance ρ and a battery V, switched "on" or "off" by a switch K, according to the voltage state of the tunneldiode.

The tunneldiode characteristic is simplified as represented in figure 3. The tunneldiode in the high voltage state (e.g. TD1) has its switch in the "a" position (V in serie with ϱ) and the tunneldiode in the low voltage state (TD2) has its switch in the "b" position (V off).

The negative trigger pulse should be equivalent to switching U "off" and "on" by means of the switch K1.

It may be calculated that r (inner impedance of the power supply and trigger generator) has no influence on the dynamic behaviour of the circuit, therefore r shall be neglected in the following mathematical treatment.

In this case (Fig. 2):

$$I_1 = \frac{U(R+\rho)}{2R\rho} - \frac{V}{2\rho} \tag{1}$$

$$I_2 = \frac{U}{2\rho} + \frac{V(R+2\rho+pL)}{2\rho(\rho+R+2pL)} \tag{2}$$

$$I_3 = \frac{U}{2\rho} - \frac{V(R+2pL)}{2\rho(R+\rho+2pL)} \tag{3}$$

If the voltage U steps now to zero ($t=0$), the currents decrease in all the branches of the circuit. I_2 decreases below I-valley (Fig. 3) and the tunneldiode (TD1) jumps down. The switch K2 will then be in the "b" state.

Both tunneldiodes are in the low voltage state. The currents I_2 and I_3 are computed now (all switches in "b" state):

$$I_{2(t\geqslant 0)} = \frac{V}{2(R+\rho)} e^{-t(R+\rho)/2L} \tag{4}$$

This current is positive, which means (Fig. 2) that the tunneldiode (TD1) is back-biased by a current which decreases with a time constant $2\,L/R+\rho$.

$$I_{3(t\geqslant 0)} = \frac{V}{2(R+\rho)} e^{-t(R+\rho)/2L} \tag{5}$$

In this case the tunneldiode (TD2) is biased in the forward direction (Fig. 3). Both currents result from the discharge of the inductivity L. When K1 is switched to the "a" state ($t=T$), and U steps back to its initial value (end of the trigger pulse) the currents in the two tunneldiodes increase and TD2, which former was in the low voltage state reaches first I-peak (because of the biasing of I_3 ($t \geqslant 0$) and jumps into the high voltage state. The switch K3 is in the "a" state and K2 stays in the "b" state.

If T is the time during which U is zero (duration of the trigger pulse), the currents in the tunneldiodes then have following expressions:

$$I_{2(t\geqslant T)} = -\frac{U}{2\rho} + \frac{VR}{2\rho(\rho+R)} + \frac{V}{2(R+\rho)} e^{-t(R+\rho)/2L}\left(1 + e^{-T(R+\rho)/2L}\right) \tag{6}$$

$$I_{3(t\geqslant T)} = \frac{U}{2\rho} - \frac{V(R+2\rho)}{2\rho(\rho+R)} + \frac{V}{2(R+\rho)} e^{-t(R+\rho)/2L}\left(1 + e^{-T(R+\rho)/2L}\right) \tag{7}$$

The situation is completely reversed, with the difference that a commutation transient has appeared which increases the current in the tunneldiode being in the high voltage state, and decreases in the same amount, the current in the diode which now is in the low voltage state. From this effect it follows that, during a certain period determined by the time constant $2\ L/R+\rho$ the circuit is less sensitive. When this period has expired, a next trigger pulse may bring the multivibrator to its initial state.

2.2 — Practical circuit design for one flip-flop

Foregoing computations are only useful for giving a qualitative idea of the behaviour of the circuit and are not efficient for real components determination. In fact the important regions of the tunneldiode characteristic around I-peak and I-valley are not linear at all. Also the assumption of equal equivalent resistances in both high and low voltage states is at least a factor of two wrong. Further the tunneldiodes were supposed to jump immediately from high to low voltage state and vice versa. The influence of the inner reactive components of the diodes was also neglected. These components (Fig. 4) give a low-pass filter characteristic which delays the input pulse action. The actual rise time of the trigger pulse causes also the tunneldiode not to jump immediately when the trigger pulse appears at the input.

The trigger amplitude was considered to be equal to U. Practically it is sufficient that the trigger pulse makes I_2 lower than I-valley.

V is a function of the used semiconductor material (Ge, Si, GaAs), ϱ depends on the tunneldiode type and peak current; U, R, r and L are to be chosen.

Since the qualitative formulas give no indications for this, a stable and sensitive bias point is to be determined first. Figure 5 represents the static input characteristic of the flip-flop measured between x and y (Fig. 1).

This curve results from the series-parallel combination of two tunneldiodes and two resistors (4) and can be calculated by simple superposition.

The bias point will be determined by the power supply impedance considered as the loadline of the system. This loadline must be drawn so that the A state is the only possible static state.

If the voltage is lowered below V_B (as for a usual tunneldiode univibrator) the circuit jumps to the low voltage state. The necessary trigger current is about $I_A - I_B$.

If the load resistance is too great, the sensitivity is low and it is possible to have both tunneldiodes in high or low voltage state. With low load resistance a great sensitivity of the tunneldiodes themselves is possible and the static state is well determined, but an important part of the trigger current flows into the power supply circuitry. Hence the load resistor r (Fig. 1) affects the total sensitivity in two conflicting ways.

For computing this influence, the characteristic region around the bias point A is considered as a straight line (Fig. 6) with angular coefficient R. The trigger current $I_t = I_A - I_B$ can now be calculated as a function of r, R, I_D, V_D, I_B and V_B (Fig. 6). The total trigger current I_T, inclusive the trigger losses in r, is:

$$I_T = I_t \frac{R + r}{r} \qquad \text{(Fig. 7)}$$

$$I_T = I_D - I_B - \frac{V_B - V_D}{r}$$

$$\frac{\delta I_T}{\delta r} = + \frac{V_B - V_D}{r^2}$$

Consequently the sensitivity increases when r decreases. The lower limit of r is determined by the effective value of the negative circuit resistance. The maximum sensitivity cannot be chosen for a stable and reliable function because of differences in input characteristics of the two stable states, due to the possible differences in the characteristics of two individual tunneldiodes. Also an asymmetry introduced by load influences of following stages will be given, the influence of which cannot completely be taken into account. Possible temperature and supply voltages changes are further limits to the sensitivity.

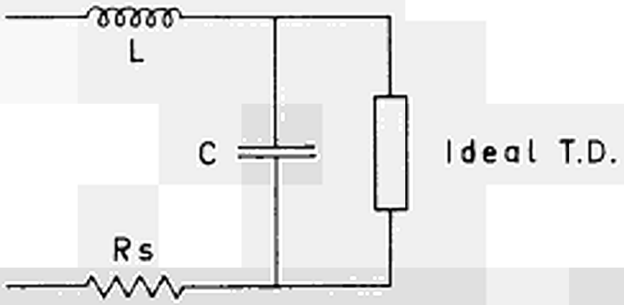

Fig. 4 — Equivalent tunneldiode circuit with components drawn in lumped form

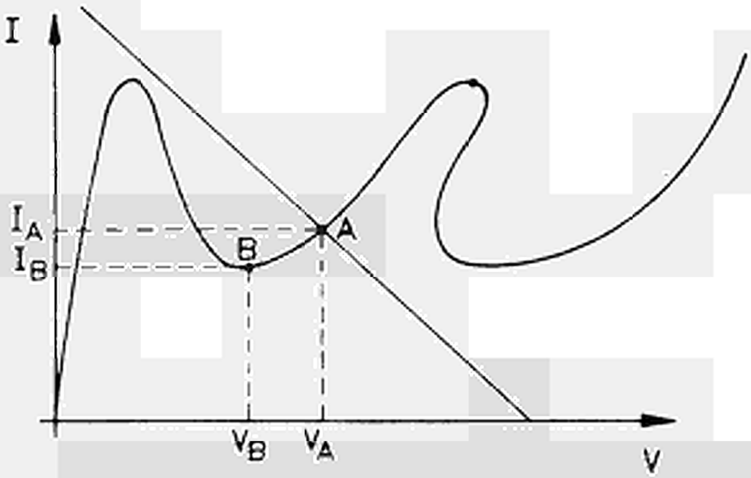

Fig. 5 — Static input characteristic of flip-flop with power supply impedance drawn as the loadline of the circuit

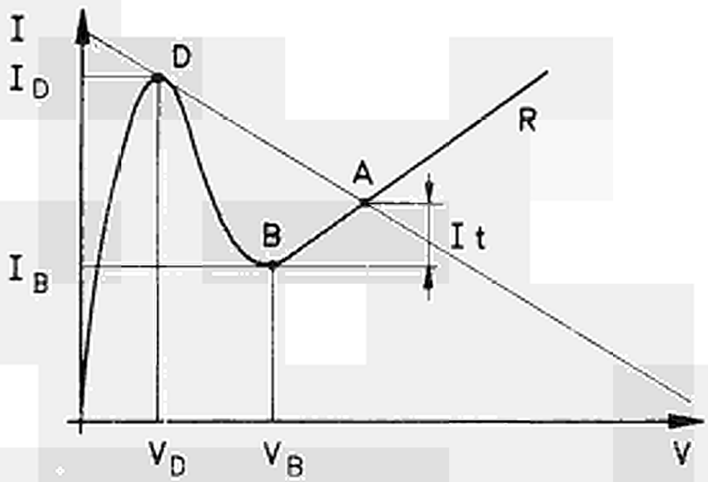

Fig. 6 — Linearisation of the static input characteristic for determination of optimal bias point

When r and U are fixed for optimal sensitivity and stability, R and L will determine the resolution of the flip-flop; this means that the time constant $2\ L/R + \varrho$ must be chosen great in comparison with the maximum pulse width of the trigger pulses (because of formulas 4, 5) and small in comparison with the shortest time between two trigger pulses (formulas 6, 7).

For a given pulse width, one can say the greater the initial memory current $i_M = V/\varrho + R$ (4, 5), the smaller could be the time constant. Hence for high counting rates and resolutions, R must be as small as possible, with a lower limit influenced again by the necessity of having effective negative resistance. When R is chosen, L is given by the necessary memory time constant.

3 — ADDRESS SCALER

3.1 — **Circuit details** (Fig. 8)

Cascading two or more of these multivibrators presents different problems as for instance undesired coupling between stages through the bias networks. Because of the high duty cycles and fast transients which are possible, only resistive filter elements can be used. But as explained before, the power supply impedance must be low. There is so to be found an optimum compromise between sensitive triggering, best biasing characteristic, small undesired coupling and reasonable power consumption.

Since a flip-flop is sensitive to both positive and negative pulses, unipolar trigger pulses must be used. The impedance of elements used to cut away the undesired polarity of pulse must be adapted to the input characteristics of the following flip-flop. Therefore ordinary diodes cannot be used; also because the signal levels are to low to give sufficient conduction in the forward direction. Backward diodes might be used, but better results were obtained with a tunnel-diode univibrator as coupling element between two stages.

This solution gives a quasi complete disappearance of the undesirable polarity of pulse and a regeneration of useful polarity; it permits also to load less the foregoing stage and to trigger the following stage with a greater trigger current.

The increase in propagation delay, with backward diodes as polarity sensitive coupling elements, is smaller than with univibrators. But the mentioned advantages of univibrators are predominating for the design.

It may be mentioned that, if a pulse of undesired polarity follows a pulse of the chosen trigger polarity at a distance small enough in relation to the memory time constant of a flip-flop no triggering action will be introduced by such pulse. This property may be exploited in certain special cases to avoid polarity sensitive coupling elements.

The spread in the tunneldiode characteristics and in the resistor values causes a spread in the characteristics of the different stages. A trimpot in serie with the filter resistor allows the adjustment of each stage to the same jump down potential (voltage for which the diode in the high voltage state jumps down). Then a greater range of supply voltage, optimum sensitivity and minimum propagation delay is possible.

3.2 — **Results**

A ten-stage address scaler was built and tested.

Counting rates:

With 8 mA, short pulses (full width at half maximum (FWHM) about 1.5 mμs) counting rates to 200 Megapulses per second (limit of the pulse generator) were tried successfully. With 8 mA peak to peak sine waves, the scaler was tested up to 300 Mc/s.

Dynamic range:

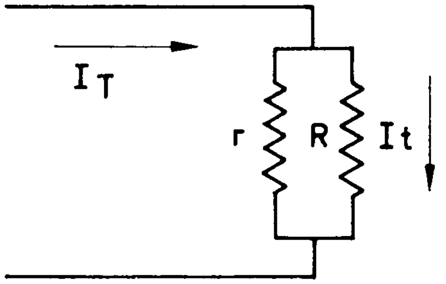

Fig. 7 — Equivalent input impedance of the flip-flop around the bias point

The limits of the dynamic range are given theoretically by the width of the pulse at the trigger level. Practically this limit is fixed by the undershoot of the trigger pulse which flops the multivibrator to its initial state. The minimum possible trigger current was about 3 mA. Pulses up to 20 mA were used successfully. (FWHM 1.5 mμs).

Voltage stability:

With 8 mA pulses $\pm$ 5% supply voltage changes were allowed for failureless function.

Temperature stability:

The scaler was tested with 8 mA pulses between -15°C and $+55^{\circ}$C and no failure was detected (counting rate: 100 Mc/s).

Propagation delay:

As explained before it was in this case, important to minimize the trigger-propagation delay and especially the jitter of this delay which influences the accuracy of time definition in time coders. The measured delay for the ten-stage scaler was about 16 mμs (i.e. 1 mμs per flip-flop and 0.6 mμs per coupling univibrator).

The jitter measured at constant voltage and temperature was less than 0.08 mμs (resolution of the coincidence unit which was used for this measurement).

Temperature coefficient of delay: -0.2‰ per 1°C.

Supply voltage dependence of delay: 2‰ for 1% voltage drift.

Because of the necessary constance of propagation delay the temperature stability of delay was improved by introducing a slightly temperature sensitive power supply voltage.

4 — PULSE DERANDOMISER

4.1 — Statistical design

The regularizing action of fast scales on random pulses was demonstrated in different reports (5). If a slow scaler is preceded by a fast scale of K, the relative counting loss in the slow scale (6) for small counting losses is given by:

$$X = Q_K\,(Nt) = 1 - \sum_{m=0}^{K-1} \frac{e^{-Nt}\,(Nt)^m}{m\,!}$$

N: actual counting rate

t : dead time of the slow scaler

These counting losses are reduced by factors much greater than the scaling factors involved if the resolution of the fast scale is good enough.

In figure 9 the dependence of counting losses from the input pulse rate is approximately given for the following conditions:

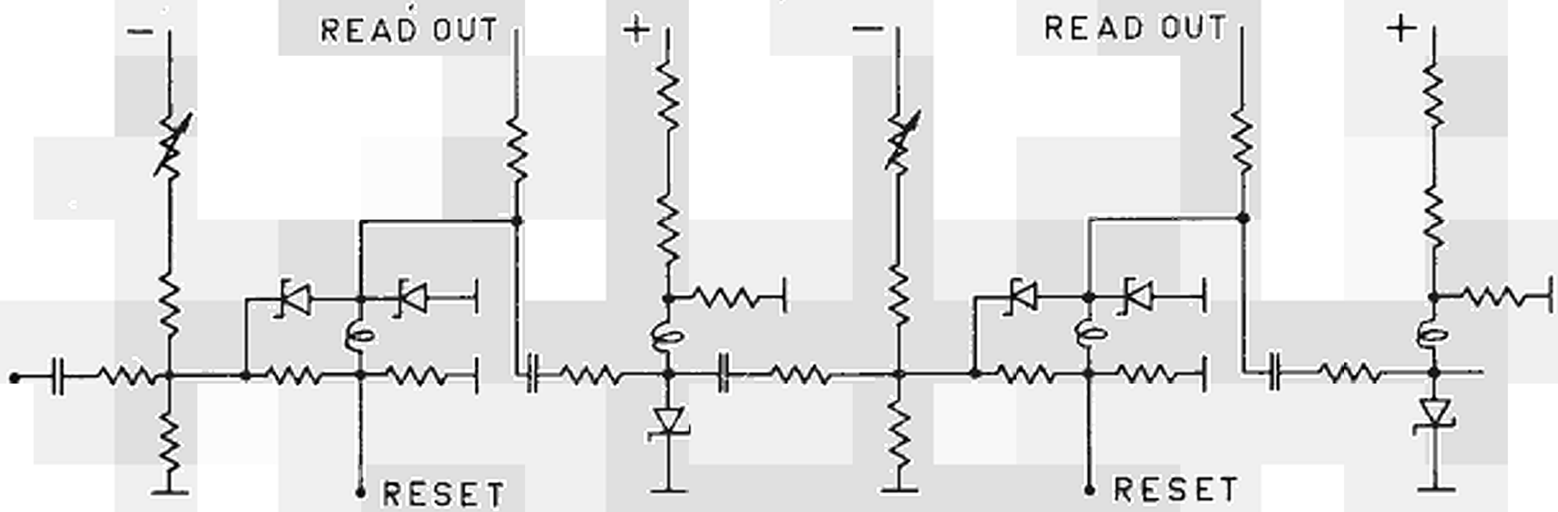

Fig. 8 — Two cascaded flip-flops with coupling univibrators

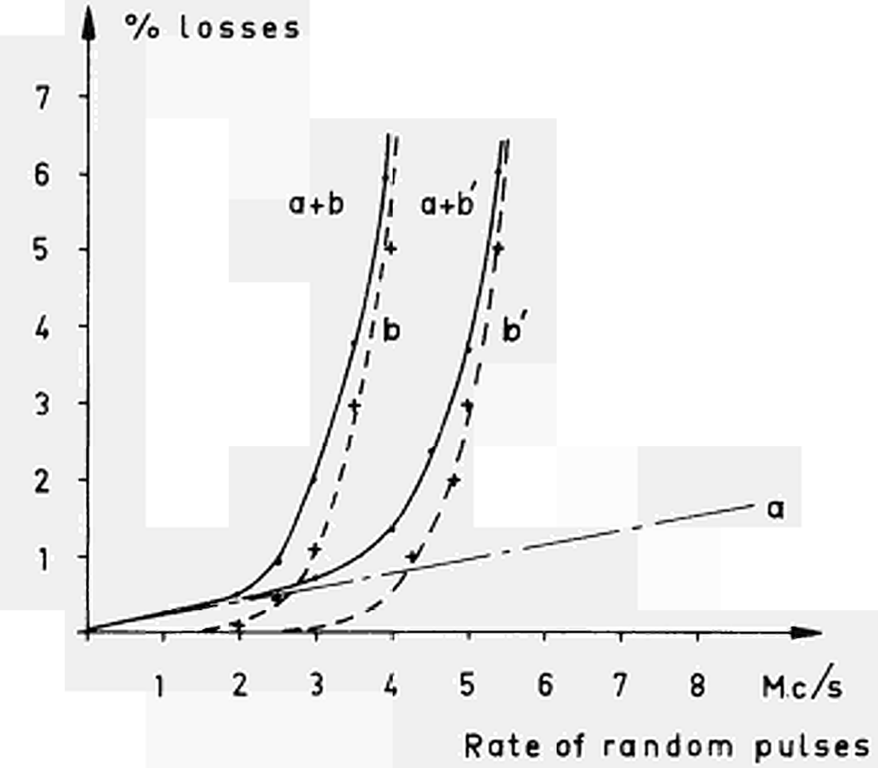

Fig. 9 — Dependence of counting losses from input pulse rate:
a) for a fast scaler with 2 mμs resolution;
b) for a slow scaler with 1 μs resolution, preceded by an ideal scale-of-eight;
b′) for the same slow scaler, preceded by an ideal scale-of-ten;
a + *b*) total losses for the derandomizer system with a scale-of-eight;
a + *b′*) total losses for the derandomizer system with a scale-of-ten.

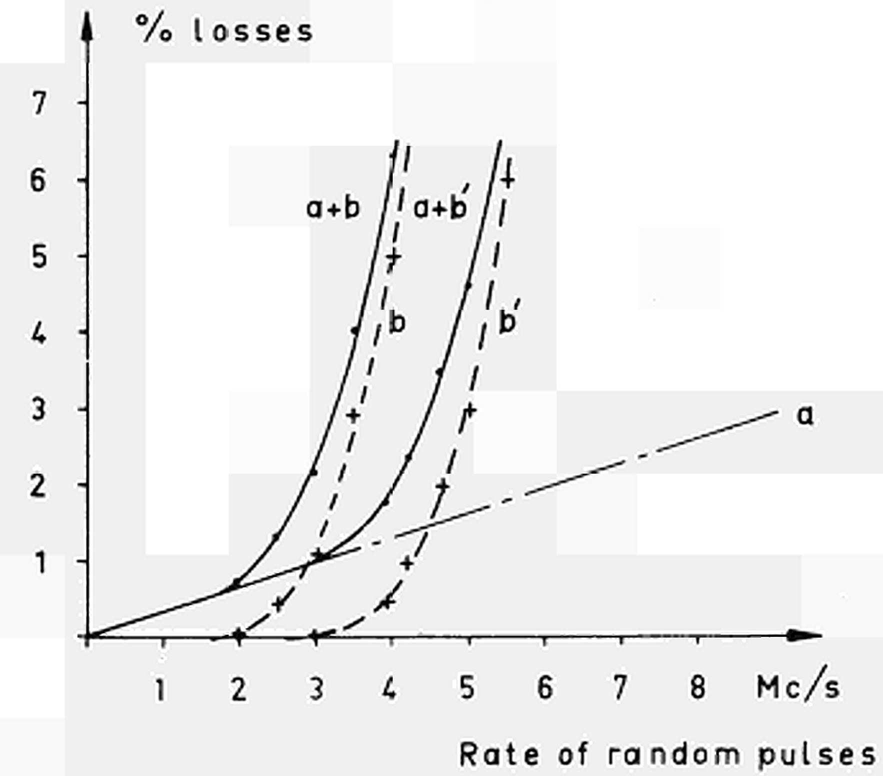

Fig. 10 — Dependence of counting losses from input pulse rate; as for figure 9 but fast scaler with 3.5 mμs resolution

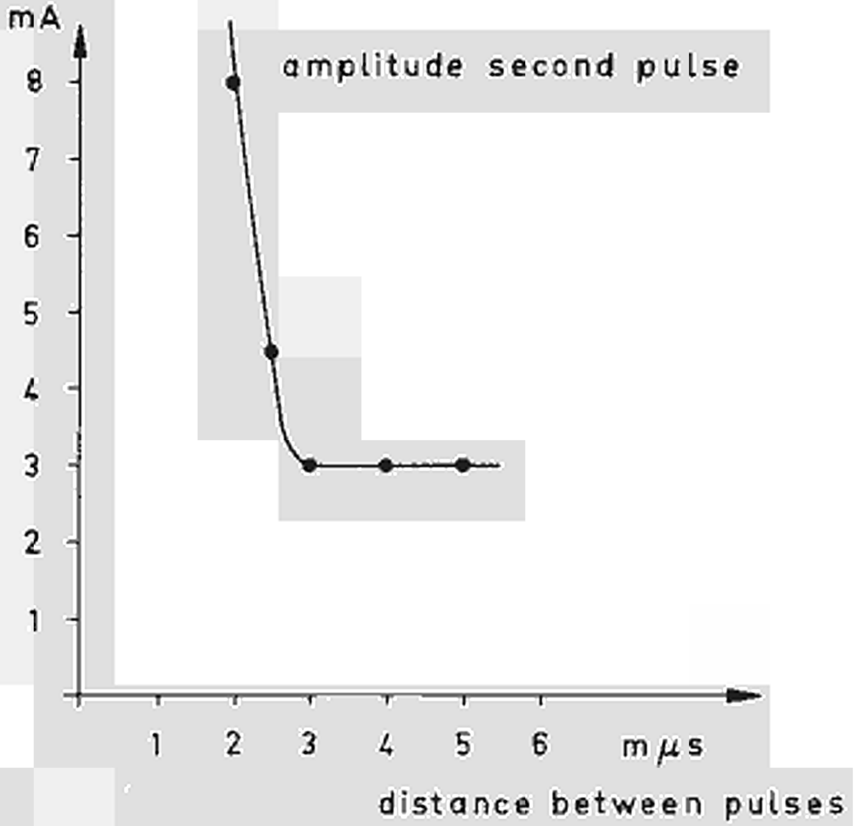

Fig. 11 — Double-pulse resolution vs. necessary trigger amplitude of the second pulse

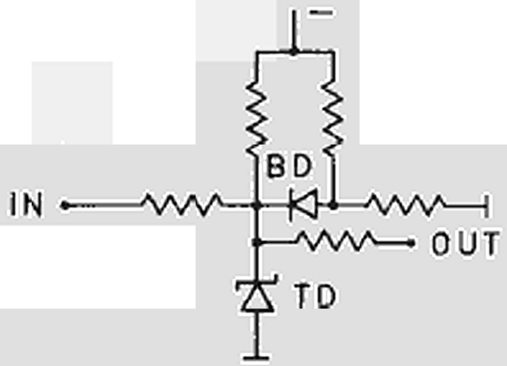

Fig. 12 — Fast tunneldiode discriminator used as input threshold

a) A derandomizer with three flip-flops (scale-of-eight) and the best double-pulse resolution reached (2 mμs);

b) A scale-of-eight with no counting losses followed by a slow scaler with 1 μsec resolution;

c) The sum of (a) and (b) which gives approximately the counting losses of the derandomizer scaler system.

It follows that for allowable counting losses of 1%, the counting rate can be 2.5 Mc/s (resp. 3.6 Mc/s for a scale-of-ten). From this results an improvement factor I in maximum pulse rate, that can be given by the expression

$$I = \frac{n_1 \text{max. (derandomiser + slow scale)}}{n_2 \text{max. (slow scaler) . scale factor}}$$

for fixed counting losses (n max. = max. rate). Optimum values for I are possible with well adapted derandomizerscaler combinations. (For the example: $I = 31$ resp. 36, i.e. a scale-of-ten will be better than a scale-of-eight).

For a double-pulse resolution of 3.5 mμs (derandomizer with threshold in the input) a scale-of-eight is more optimal as can be seen from figure 10 ($I = 28.7$ resp. 26).

4.2 — Double-pulse resolution

The best double-pulse resolution of 2 mμsec was measured with 8 mA clipped pulses (rise time: 0.5 mμs; full width at half maximum (FWHM): 0.8 mμs).

Figure 11 gives the minimum triggering amplitude of the second pulse, in function of the delay between the two pulses. The first pulse was kept to a constant value of 3 mA (3 mA = minimum trigger value). After 3 mμs all the influences of the preceding switching seem to have vanished.

It was noted that, if a third pulse follows a 2 mμs spaced twin pulse (amplitude 8 mA) the third pulse must follow at least 6 mμs after the second one for accurate triggering. This increases theoretically the counting losses. The amount of this loss however is negligible in comparison with the pulse losses due to pulses occurring during the normal 2 mμs dead time of the flip-flop; e.g. with 2 Mc/s random pulses is the ratio about 10^4.

4.3 — Pulse shape influence

For the moment good pulse sources to utilize the maximum performance of this circuit are not available. The practical limits are mostly given by the trigger pulse shape.

On account of computation of this shape influence on the real resolution of the circuit, two principles should be observed:

The pulse width at the trigger level must be less than 5 mμs; if not, the stored information is lost. Enlarging L may increase this maximum storage time but reduces the resolution;

Resolution must be measured on trigger level. Pile-up effects make the pulses merge in each other, and on trigger level the pulses are much nearer than on peak level. (Distance between trailing edge of the first pulse and leading edge of the second one, instead of peak to peak distance). The pulse form is also partially responsible for the sharp rising slope of the curve in figure 10 (distances measured from peak to peak).

Because of this predominating influence of pulse shape, no absolute double-pulse resolution definition can be given. For each case the pulse shape influence must be taken into account separately, i.e. the multivibrator has a theoretical possible resolution better than practical pulses will allow. This resolution is not very much higher than the switching time of the tunneldiodes (smaller than one nanosecond today).

4.4 — Input discriminator

The first binary can itself serve as a discriminator. But because of the possible differences between the two stable states, the accuracy of the discriminating action is not very good. So in most cases a good foregoing discriminator should be useful. Such unit could also avoid spurious triggering by pulses of wrong polarity, and undershoot effects. A higher sensitivity is possible and the flip-flop may be triggered with narrow, fast pulses of about constant amplitude if a tunneldiode univibrator is taken for discrimination.

However the double-pulse resolution of such discriminator is normally worse than that of the flip-flop, and the resolution of the whole system is reduced.

With the circuit of figure 12 the obtained double-pulse resolution was 3.5 mμs, but the sensitivity is increased by a factor five at least.

This resolution can still be improved by the use of low-inductivity backward diodes and special circuit lay-out.

5 — CONCLUSION

The limits of tunneldiode multivibrators are not yet reached. The use of faster tunneldiodes, higher peak currents, practically lower inductivities and, also inforced, narrower pulses could allow much higher counting rates.

An important improvement of the circuit would be the use of a non-linear memory element to sharpen the exponential drop of the memory current.

For very short propagation delays, a gated tunneldiode counter should allow delays around 1 mμs; but the bipolar characteristic of the tunneldiode may cause some difficulties.

Anyhow for much faster circuitry good lumped elements, to use with the low impedance levels of fast tunneldiodes, will be difficult to realize and power consumption risks also to rise quickly.

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